OPTIMUM FIELD THEORY DESIGN OF STEPPED E-PLANE FINNED WAVEGUIDE TRANSFORMERS OF DIFFERENT INNER CROSS-SECTIONS

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ABSTRACT

This paper presents a new design of broadband double-stepped ridged and all-metal finned waveguide transformers which combines the advantage of constant ridge thickness with that of the optimum matching potential of different waveguide inner cross-section dimensions. Based on the modal scattering matrix method, the rigorous design takes into account the influences of both, the finite ridge or fin thicknesses and the higher-order mode interaction at all discontinuities. Computer optimized design data are given for compact transformers for WR112 and WR42 input waveguides achieving a return loss of more than 33dB for the whole waveguide bands. The theory is verified by measurements.

INTRODUCTION

All-metal finned waveguide circuits have found many applications in microwave and millimeter devices, such as components for converters, attenuators, matching networks, mixers, or as resonators in evanescent-mode lowpass and bandpass filters, [1] - [10]. The advantages of these circuits include large dominant-mode bandwidth, low characteristic impedances, and the possibility of low-cost, low-loss E-plane integrated circuit designs. Of particular importance, therefore, are the transitions between ridged or finned waveguides of different gap height and waveguide inner cross-sections as well as the junction to the rectangular input waveguide. This paper describes an exact field theory design of broadband and compact ridged and finned waveguide transformers with optimum stepped transitions for both the ridges and the different outer cross-sections from the input waveguide to the smaller housing of the ridged waveguide (Fig. 1).

Based on Cohn's transmission-line approximation [11], Hensperger [2] presented a transformer from a WR-112 rectangular to the Airtron ARA-133 double-ridged waveguide [12]. That design includes a continuous transition from the input waveguide to the smaller ridged waveguide cross-section. The main disadvantage is, however, that a constant ratio (of 0.25) in ridge thickness to waveguide width has been utilized which requires also the ridge thickness to be tapered along the whole length of the transition. Therefore, the design is considered to be not very appropriate for the modern fabrication methods for microwave and millimeter-wave E-plane

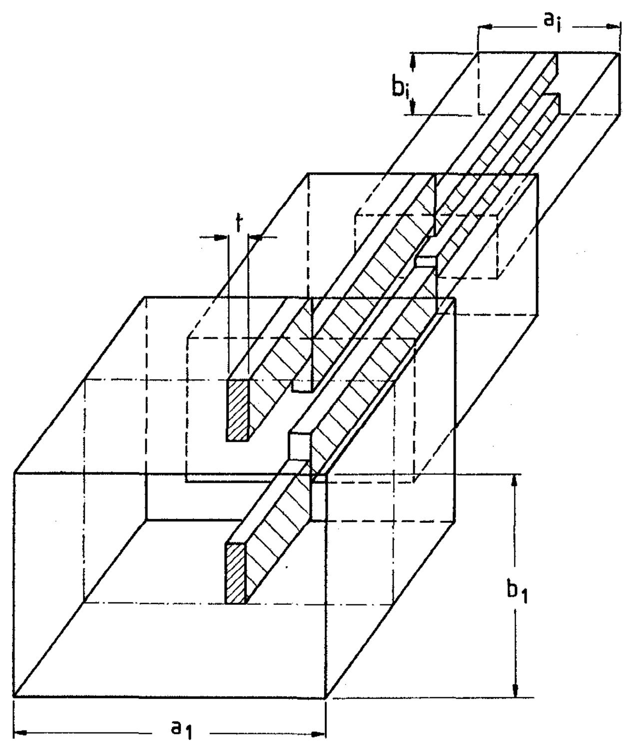

Fig. 1: Stepped E-plane all-metal finned waveguide transformer of different inner cross-sections

integrated circuits, such as etching or computer controlled milling techniques. A stepped transformer with constant ridge thickness has been presented in [8], but the transformer housing dimensions are kept unchanged. Hence, the advantage of the high bandwidth potential inherent in a proper choice of the waveguide cross-sections is not utilized. Moreover, with the exception of [4], [8], most of the techniques presented for the analysis of all-metal finline discontinuities neglect the influence of the finite metallization thickness which has turned out to be important for the design of E-plane integrated circuits.

The theory given in this paper, which includes higher order mode coupling effects as well as the finite ridge or metal fin width, is based on modal field expansion into orthogonal eigenmodes [8], [10]. An optimization procedure based on a modified direct search method, the evolution strategy, leads to optimum stepped transformer designs including the waveguide inner cross-sections. This design combines the advantage of stepped ridges of constant thickness, well suited to the printed E-plane technology, with that of the additional matching potential of the different cross-section dimensions. Moreover, the design achieves an improved return loss behaviour and a compact structure. The design data given may be transferred into other waveguide bands by suitable frequency scaling relations which include the metal fin thicknesses. Measured results verify the theory given.

THEORY

For the computer-aided design of the all-metal E-plane finned waveguide transformer (Fig.1), the modal S-matrix method [8], [10] is applied. The structure is decomposed into two key building blocks: double-plane waveveguide step [10], and the finite-length ridge section (Fig.2). Combination with the modal scattering matrices of the corresponding intermediate homogeneous waveguide sections of finite or zero lengths, respectively, yields the total scattering matrix of the transformer.

The electromagnetic field in the subregions i = 0, I, II (Fig. 2)

$$\vec{E}^i = \nabla \times (A^i_{Hz}\vec{e}_z) + \frac{1}{j\omega\epsilon}\nabla \times \nabla \times (A^i_{Ez}\vec{e}_z) \qquad (1)$$

$$\vec{H}^i = \nabla \times (A^i_{Ez}\vec{e}_z) - \frac{1}{j\omega\mu}\nabla \times \nabla \times (A^i_{Hz}\vec{e}_z)$$

is derived from the vector potentials

$$\vec{A}^i_{Hz} = \sum_{q=1}^{\infty} (\sqrt{Z^i_{Hq}}) \cdot T^i_{Hq}(x,y) \cdot [V^i_{Hq}\exp(-jk^i_{zHq}z) + R^i_{Hq}\exp(+jk_{zHq}z)] \qquad (2a)$$

$$\vec{A}^i_{Ez} = \sum_{p=1}^{\infty} (\sqrt{Y^i_{Ep}}) \cdot T^i_{Ep}(x,y) \cdot [V^i_{Ep}\exp(-jk^i_{zEp}z) - R^i_{Ep}\exp(+jk_{zEp}z)], \qquad (2b)$$

with the wave impedances

$$Z^i_{Hq} = (\omega\mu_0)/(k^i_{zHq}) = 1/Y^i_{Hq},$$
$$Y^i_{Ep} = (\omega\epsilon_0)/(k^i_{zEp}) = 1/Z^i_{Ep}\,. \qquad (3)$$

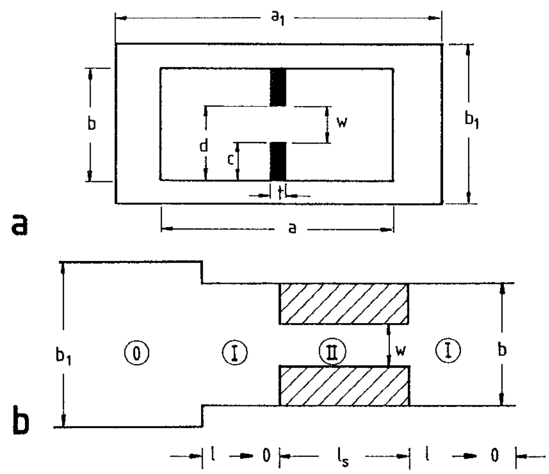

Fig. 2: Single transformer section
a) Cross-sectional dimensions
b) Longitudinal section dimensions

$V^i_{H,E}$, $R^i_{H,E}$ are the TE- and TM-mode wave amplitudes of the forward and backward waves, respectively, which have to be related to each other at the corresponding discontinuity. This will yield the corresponding scattering matrix relations. k_z are the propagation factors, and T^i_{Hq}, T^i_{Ep} are the cross-section eigenfunctions of the corresponding waveguide structures under consideration [8], [10].

For the E-plane finned or ridged waveguide eigenvalue problem (Fig.2, region II), the transverse resonance method is used [8], [10]. This procedure reduces the size of the characteristic matrix equation to a quarter of the original size. Furthermore, it makes the method very flexible because an arbitrary number of subregions (e.g. fin-lines with multiple inserts) may be easily taken into account in the matrix system, simply by multiplying the additional transmission-line matrices of the corresponding insert subregions. Moreover, there are no poles in the determinant function of the resulting characteristic matrix equation.

Matching the tangential field components of the regions involved at the common interface yields the modal scattering matrix (S) of the related discontinuity where the submatrices are already described in [8]. The series of step discontinuities, for a complete transformer structure, is calculated by direct combination of the single modal scattering matrices [8].

The computer-aided design is carried out step by step, in order to reduce the requirements for the optimization process: First, the transitions of the waveguide inner dimensions are assumed to be linearly stepped. Second, initial transformer dimensions are calculated using Cohn's method [11]. Finally, an optimization program applying the evolution strategy method is used; an error function is minimized with respect to a parameter vector, which contains the slot widths and section lengths, until a specified return loss value (e.g., 30 dB) is achieved over the desired frequency range.

The number of TE-modes, TM-modes, and cross-sectional expansion terms [8] are chosen to be 5, 2, 8 for the optimization, and 12, 7, 8 for the final analysis, respectively. The numbers have been obtained by checking the convergence behaviour against measurements [8], cf. Fig. 3. Furthermore, the dashed curve (5, 2, 8) in comparison with the solid curve (12, 7, 8) in Fig. 3 demonstrates the significant influence of the higher order modes on the correct prediction of the return loss peaks which may be particularly important for narrow-band designs.

RESULTS

Fig. 4 shows the comparison between Hensperger's [2] and our design, at the example of the five-step WR-112 (7.05 - 10 GHz) waveguide to ARA-133 ridged waveguide transformer [12]. Additional to the advantage of constant ridge thickness, the double-stepped transformer design achieves a reduction of the maximum VSWR within the whole frequency range; moreover, the overall transformer length is slightly decreased from about 48 to 46 mm. The minimum return loss is better than 36 dB for the optimized structure which may demonstrate the advantage of the computer-aided field theory design.

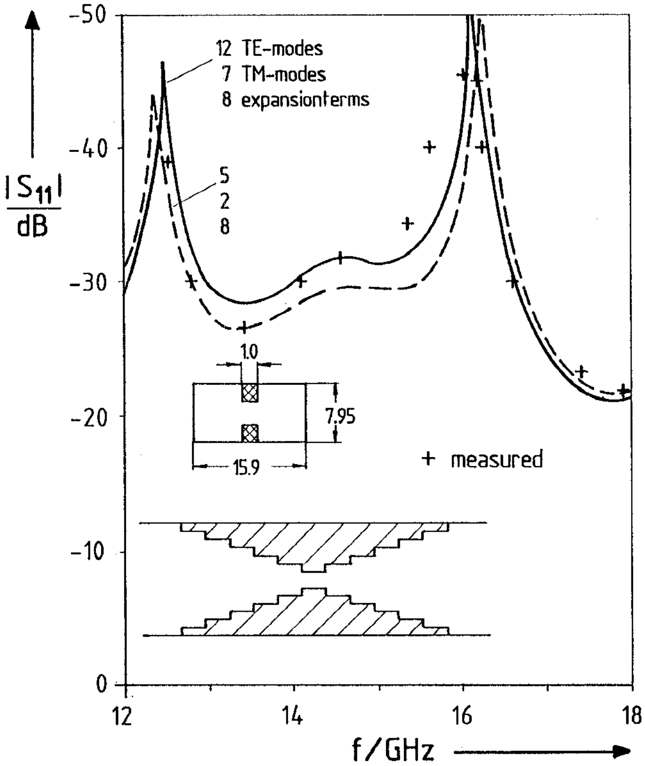

Fig. 3: Comparison between measurements (+ + +) [8] and theory.
Input reflection coefficient $|S_{11}|$ in decibels as a function of frequency of an optimum stepped Ku-band (12 - 18 GHz) ridged waveguide transformer with two symmetrical five-step sections and with constant waveguide inner cross-section dimensions.

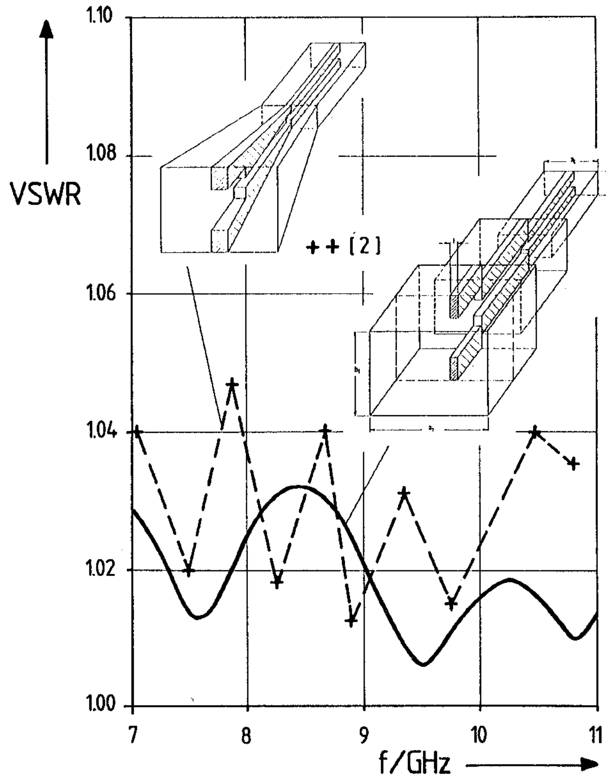

Fig. 4: Optimum five-step WR-112 (7.05 - 10 GHz) waveguide to ARA-133 ridged waveguide transformer
Input VSWR as a function of frequency of the optimum design (solid line) in comparison with Hensperger's linearly tapered ridge thickness design (dashed line); t = 6.5 mm.

In Fig. 5, the result of a frequency-scaled version of the optimized configuration of Fig. 4 is presented for a WR-42 input waveguide (K-band: 18 - 26 GHz). The good return loss behaviour is maintained beyond the limit of the waveguide band of the input waveguide. The slight reduction in return loss (minimum value 33 dB), as compared with the design of Fig. 4, is due to the fact that a modified ratio of the ridge thickness to waveguide width, t/a, has been chosen in order to obtain an adequately manufacturable ridge thickness (t = 2.4mm).

CONCLUSION

The rigorous modal S-matrix method presented achieves the exact computer-aided design of new broadband double-stepped ridged and all-metal E-plane finned waveguide transformers which combines the advantage of constant ridge thickness, well suited to the printed E-plane technology, with that of the additional matching potential of different inner waveguide dimensions. Application of a modified direct search method leads to optimum low input VSWR over the whole waveguide band and to compact designs, as has been demonstrated for WR-112 and WR-42 metal finline transformer examples. The design data are transferable into other frequency bands of interest. Since the theory includes the finite thickness of the fins as well as the higher order mode interaction at all discontinuities, measurements verify the theory by good agreement.

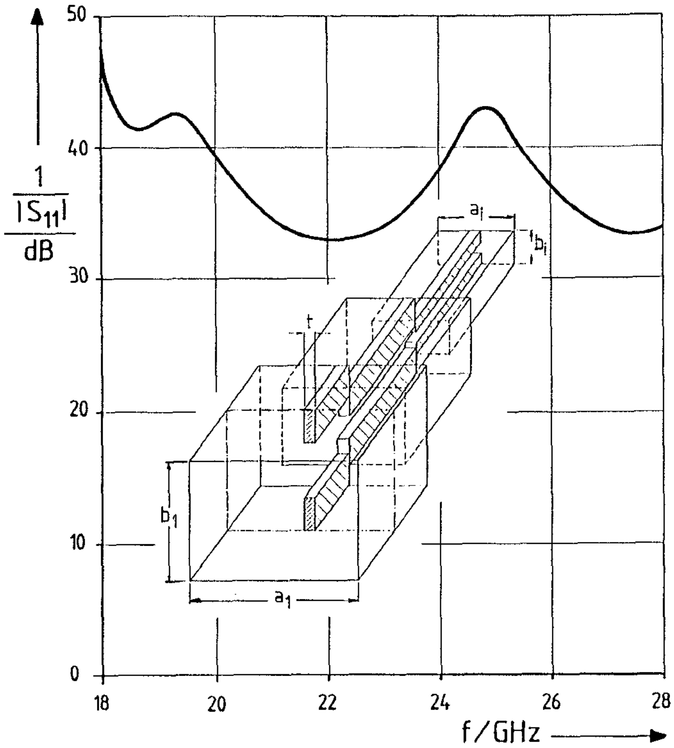

Fig. 5: Optimum five-step WR-42 (18 - 26.5 GHz) waveguide to E-plane all-metal finned waveguide transformer
t= 2.4 mm, finned waveguide: a_6 = 9.745mm, b_6 = 4.127mm, w = 1.66mm. Frequency-scaled version of the optimized configuration of Fig. 4. Input reflection coefficient $|S_{11}|$ in decibels as a function of frequency.

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